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(54) Title: COMBINED INTERFERENCE CANCELLATION AND MAXIMUM LIKELIHOOD DECODING OF SPACE-TIME BLOCK CODES

(57) Abstract

Block-encoded transmissions of a multi-antenna terminal unit are effectively detected in the presence of co-channel interfering transmissions from other multi-antenna terminal units, when the base station has a plurality of antennas, and interference cancellation is combined with maximum likehood decoding. The signals received in one base station antenna are employed in processing the signals received in a second base station antenna so as to cancel the signals of one terminal unit, while decoding the signals transmitted by the other terminal unit. Zero-forcing and MMSE approaches are presented. In another embodiment of the invention, the basic decoding approach is used to obtain an initial estimate for the symbols from each terminal. Assuming that the signals of the first terminal unit has been decoded correctly, the receiver employs this initial decoded signal of the first terminal unit to cancel their contribution to the signals received at the base station antennas while decoding the signals of the second terminal unit. This process is then repeated assuming that the signals of the second terminal unit have been decoded correctly; the receiver employs this initial decoded signal of the second terminal unit to cancel their contribution to the signals received at the base station antennas while decoding the signals of the first terminal unit. The above disclosed techniques are variable for any number K of terminal units concurrently transmitting over a given channel, where each terminal unit is using a space-time block code with N transmit antennas, and a base station has at least K receive antennas.

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Combined Interference Cancellation and Maximum Likelihood Decoding of Space-Time Block Codes

Reference to Related Applications

This application claims the benefit of U.S. Provisional Application No. 60/061,145, filed October 6, 1997. This application is also related to U.S. Application 09/074224, filed May 7, 1998, titled "Transmitter Diversity Technique for Wireless Communications".

Background of the Invention

This invention relates to wireless communication and, more particularly, to techniques for effective wireless communication in the presence of fading, co-channel interference, and other degradations.

Rapid growth in mobile computing and other wireless data services is inspiring many proposals for high speed data services in the range of 64-144 kbps for micro cellular wide area and high mobility applications and up to 2 Mbps for indoor applications. Research challenges include the development of efficient coding and modulation, and signal processing techniques to improve the quality and spectral efficiency of wireless communications and better techniques for sharing the limited spectrum among different high capacity users.

The physical limitation of the wireless channel presents a fundamental technical challenge for reliable communications. The channel is susceptible to time-varying noise, interference, and multipaths. Power and size limitations of the communications and computing device in a mobile handset constitute another major design consideration. Most personal communications and wireless services portables are meant to be carried in a briefcase and/or pocket and must, therefore, be small and lightweight. This translates to a low power requirement since small batteries must be used.

However, many of the signal processing techniques which may be used for reliable communications and efficient spectral utilization demand significant processing power, precluding the use of low power devices. Continuing advances in VLSI and integrated circuit technology for low power applications will provide a partial solution to this problem. Still, placing most of the signal processing burden on fixed locations (base stations) with relatively larger power resources than the mobile units will, likely, continue to be the trend in wireless systems design.

Perhaps the single most important parameter in providing reliable communications over wireless channels is diversity. Diversity techniques which may be used include time, frequency, and space diversity

- *Time diversity:* channel coding in combination with limited interleaving is used to provide time diversity. However, while channel coding is extremely effective in fast fading environments (high mobility), it offers very little protection under slow fading (low mobility) unless significant interleaving delays can be tolerated.
- *Frequency diversity:* the fact that signals transmitted over different frequencies induce different multipath structures and independent fading is. However, when the multipath delay spread is small compared to the symbol period, frequency diversity is not helpful.
- *Space diversity:* the receiver/transmitter uses multiple antennas that are separated for reception/transmission and/or differently polarized antennas to create independent fading channels. Currently, multiple antennas at base-stations are used for receive diversity at the base. However, it is difficult to have more than two antennas at the mobile unit due to the size and cost of multiple chains of RF conversion.

Previous work on transmit diversity can be classified into three broad categories: schemes using feedback, schemes with feedforward or training information but no feedback, and blind schemes. The third category (blind schemes) relies primarily on multiple transmit antennas combined with channel coding to provide diversity. An example of this approach is

disclosed in the aforementioned copending application 09/074224, 1998, titled "Transmitter Diversity Technique for Wireless Communications," filed May 7.

Summary of the Invention

Improved performance is attained in an illustrative arrangement where K synchronized terminal units that transmit on N antennas to a base station having $M \geq K$ antennas, by combining interference cancellation (IC) and maximum likelihood (ML) decoding. More specifically, space-time block coding is employed in transmitters that employ N transmit antennas each, and the signals are received in a receiver that employs M receiving antennas. In accordance with the processing disclosed herein, by exploiting the structure of the space-time block code, K-1 interfering transmitting units are cancelled at the receiver, regardless of the number of transmitting antennas, N, when decoding the signals transmitted by a given mobile unit. In another embodiment of the principles of this invention, signals of a first terminal unit are decoded first, and the resulting decoded signals are employed to cancel their contribution to the signals received at the base station antennas while decoding the signals of the remaining K-1 terminal units. The process is repeated among the remaining K-1 terminal units. That is, among the remaining K-1, signals of a first terminal unit is decoded first and the resulting decoded signals are employed to cancel their contribution to the signals received at the base station antennas while decoding the signals of the remaining K-2 terminal units, and so on. This procedure is repeated M times, each time starting with decoding signals of a particular terminal unit. This successive procedure will yield additional performance improvement.

Both zero-forcing (ZF) and minimum mean-squared error (MMSE) interference cancellation (IC) and maximum likelihood (ML) techniques are disclosed.

Brief Description of the Drawing

FIG. 1 depicts an arrangement that, illustratively, includes a receiving base station (20) and two transmitting terminal units (10 and 30).

Detailed Description

FIG. 1 illustrates two transmitting units and one receiving unit that comport with the principles disclosed herein. However, this is merely illustrative, and the disclosed method is useful for more than two terminals ($K > 2$).

Single Transmitting Unit

Transmitting unit 10 may correspond to the transmitting circuitry in a terminal unit, while receiving unit 20 may correspond to the receiving circuitry in a base station. Terminal unit 30 is shown identical to terminal unit 10. It should be understood, of course, that each terminal unit has a receiving circuit, and the base station has a transmitting circuit. The terminal units are shown to have two antennas each. Receiving unit 20 is also shown to have two receiving antennas. Here, too, it should be kept in mind that, generally, any number, $M \geq 2$, of receiving antennas can be had. Particular advantage is realized when $M \geq K$. Since the mathematical treatment below is couched in general matrix notations, the expressions are valid for any number K and/or M.

Considering terminal unit 10, the information source provides input symbols to element 13 which develops a block code. The symbols are divided into groups of two symbols each, and at a given symbol period, the two symbols in each group $\{c_1, c_2\}$ are transmitted simultaneously from the two antennas. The signal transmitted from antenna 11 is c_1 and the signal transmitted from antenna 12 is c_2. In the next symbol period, the signal $-c_2*$ is transmitted from antenna 11 and the signal c_1* is transmitted from antenna 12. The symbols are modulated prior to transmission with constellation mappers 14 and 15, followed by pulse shapers 16 and 17, respectively, in a conventional manner.

In receiver 20, signals are received by antennas 21 and 22 and are applied to detector 25.

In the mathematical development of the algorithms disclosed herein, it is assumed that the channel from each of the two transmit antennas remains fixed over two consecutive symbol periods. That is,

$$h_i(nT) = h_i((n+1)T), \quad i = 1,2. \tag{1}$$

To ascertain the channel characteristics, the transmitter carries out a calibration session, during which pilot signals or tones are transmitted. The signals received during the calibration session are applied to channel estimator circuits 23 and 24, which are well known circuits, and the channel characteristics are thus obtained.

When only transmitter 10 is considered, the received signals at antenna 21 can be expressed as

$$r_{11} = h_{11}c_1 + h_{12}c_2 + \eta_1 \tag{2}$$

$$r_{12} = -h_{11}c_2^* + h_{12}c_1^* + \eta_2 \tag{3}$$

where r_{11} and r_{12} are the received signals over two consecutive symbol periods, h_{11} denotes the fading channel between transmit antenna 11 and receive antenna 21, h_{12} denotes channel between transmit antenna 12 and receive antenna 21, and η_1 and η_2 are noise terms, which are assumed to be complex Gaussian random variables with zero mean and power spectral density $N_0/2$ per dimension. Defining the vectors $\mathbf{r} = [r_{11} r_{12}{}^*]^T$, $\mathbf{c} = [c_1 c_2]^T$, and $\eta = [\eta_1 \eta_2{}^*]^T$, equations (2) and (3) can be rewritten in a matrix form as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \eta \tag{4}$$

where the channel matrix $\mathbf{H}$ is defined as

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}{}^* & -h_{11}{}^* \end{bmatrix}. \tag{5}$$

The vector η is a complex Gaussian random vector with zero mean and covariance $N_0 \cdot \mathbf{I}$. Defining C as the set of all possible symbol-pairs $\mathbf{c} = \{c_1, c_2\}$, and assuming that all symbol pairs are equi-probable, it can be

easily shown that the optimum maximum likelihood (ML) decoder selects from C the symbol-pair $\hat{\mathbf{c}}$ that minimizes the expression $\|\mathbf{r}-\mathbf{H}\cdot\hat{\mathbf{c}}\|^2$. This can be written as

$$\hat{\mathbf{c}} = \arg\min_{\hat{\mathbf{c}}\in\mathbf{C}}\|\mathbf{r}-\mathbf{H}\cdot\hat{\mathbf{c}}\|^2. \tag{6}$$

It was shown by S. Alamouti in "Space Block Coding: A simple Transmitter Diversity Scheme for Wireless Communications," submitted to IEEE JSAC, September 1997 that the diversity order of the above space-time block code is equivalent to that of a two branch maximal ratio receive combining (MRRC). Because of the orthogonality of the matrix **H**, Alamouti also showed that this decoding rule decomposed into two separate decoding rules for c_1 and c_2. The *uncertainty*, Δ_c, of the decoded symbols $\hat{\mathbf{c}}$ is defined as

$$\Delta_c = \|\mathbf{r}-\mathbf{H}\cdot\hat{\mathbf{c}}\|^2. \tag{7}$$

It should be noted that the above analysis addresses only receive antenna 21. When receiver 20 uses both antennas, i.e., antennas 21 and 22, two received signal vectors $\mathbf{r}_1$ and $\mathbf{r}_2$ can be defined for antenna 21 and 22, respectively, as

$$\mathbf{r}_1 = \mathbf{H}_1\cdot\mathbf{c}+\eta_1 \tag{8}$$

$$\mathbf{r}_2 = \mathbf{H}_2\cdot\mathbf{c}+\eta_2 \tag{9}$$

where $\mathbf{H}_1$ and $\mathbf{H}_2$ are the channel matrices to receive antennas 21 and 22, respectively, and η_1 and η_2 are the corresponding noise vectors. That is,

$$\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}* & -h_{11}* \end{bmatrix}, \text{ and } \mathbf{H}_2 = \begin{bmatrix} h_{21} & h_{22} \\ h_{22}* & -h_{21}* \end{bmatrix}, \tag{9a}$$

where h_{21} denotes the channel between transmit antenna 12 and receive antenna 22, and h_{22} denotes the channel between transmit antenna 11 and receive antenna 22. In this case, the ML decoding rule is

$$\hat{\mathbf{c}} = \arg\min_{\hat{\mathbf{c}}\in\mathbf{C}}(\|\mathbf{r}_1-\mathbf{H}_1\cdot\hat{\mathbf{c}}\|^2+\|\mathbf{r}_2-\mathbf{H}_2\cdot\hat{\mathbf{c}}\|^2), \tag{10}$$

and the uncertainty of the decoded symbols is defined as

$$\Delta_c = \|\mathbf{r}_1 - \mathbf{H}_1 \cdot \hat{\mathbf{c}}\|^2 + \|\mathbf{r}_2 - \mathbf{H}_2 \cdot \hat{\mathbf{c}}\|^2 . \tag{11}$$

As before, both the matrices $\mathbf{H}_1$ and $\mathbf{H}_2$ are orthogonal matrices and hence the above decoding rule also decomposes to two separate decoding rules for c_1 and c_2. *Note that the rate of transmission (of information symbols) in the space-time block coding scheme is 1*

<u>Interference Cancellation and ML Decoding: BASIC CASE</u>

FIG. 1, however, shows two terminal units, and the issue that needs to be addressed is the detection performance at the base station receiver when the two terminal units transmit over the same time and frequency channel.

In the notation below, g_{11} denotes the fading channel between transmit antenna 31 and receive antenna 21, g_{12} denotes the channel between antenna 31 and antenna 21, g_{21} denotes the channel between antenna 32 and antenna 22, and g_{22} denotes the channel between antenna 32 and antenna 22. Also, $\{c_1, c_2\}$ and $\{s_1, s_2\}$ denote the two symbols transmitted from terminal units 10 and 30, respectively.

At receiver 20, the received signals over two consecutive symbol periods at receive antenna 21, r_{11} and r_{12}, are

$$r_{11} = h_{11}c_1 + h_{12}c_2 + g_{11}s_1 + g_{12}s_2 + \eta_{11} \tag{12}$$

$$r_{12} = -h_{11}c_2^* + h_{12}c_1^* - g_{11}s_2^* + g_{12}s_1^* + \eta_{12} \tag{13}$$

Defining $\mathbf{r}_1 = [r_{11} r_{12}{}^*]^T$, $\mathbf{c} = [c_1\ c_2]^T$, $\mathbf{s} = [s_1\ s_2]^T$, and $\mathbf{n}_1 = [\eta_{11}\ \eta_{12}^*]^T$ equations (12) and (13) can be rewritten in matrix form as

$$\mathbf{r}_1 = \mathbf{H}_1 \cdot \mathbf{c} + \mathbf{G}_1 \cdot \mathbf{s} + \mathbf{n}_1 \tag{14}$$

where the channel matrices $\mathbf{H}_1$ and $\mathbf{G}_1$ between the transmitter units 10 and 30 and receive antenna 21 are given by

$$\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix}, \text{ and } \mathbf{G}_1 = \begin{bmatrix} g_{11} & g_{12} \\ g_{12}^* & -g_{11}^* \end{bmatrix} \tag{15}$$

The vector $\mathbf{n}_1$ is a complex Gaussian random vector with zero mean and covariance $N_0 \cdot \mathbf{I}$. Similarly, the received signals over two consecutive symbol periods at receive antenna 22, r_{21} and r_{22} are

$$r_{21} = h_{21}c_1 + h_{22}c_2 + g_{21}s_1 + g_{22}s_2 + \eta_{21} \tag{16}$$

$$r_{22} = -h_{21}c_2^* + h_{22}c_1^* - g_{21}s_2^* + g_{22}s_1^* + \eta_{22} \tag{17}$$

In a similar fashion, defining $\mathbf{r}_2 = [r_{21} r_{22}{}^*]^T$ and $\mathbf{n}_2 = [\eta_{21}\eta_{22}{}^*]^T$ equations (16) and (17) can be rewritten as

$$\mathbf{r}_2 = \mathbf{H}_2 \cdot \mathbf{c} + \mathbf{G}_2 \cdot \mathbf{s} + \mathbf{n}_2 \tag{18}$$

where the channel matrices $\mathbf{H}_2$ and $\mathbf{G}_2$ between transmitter units 10 and 30 and antenna 22 are given by

$$\mathbf{H}_2 = \begin{bmatrix} h_{21} & h_{22} \\ h_{22}^* & -h_{21}^* \end{bmatrix}, \text{ and } \mathbf{G}_2 = \begin{bmatrix} g_{21} & g_{22} \\ g_{22}^* & -g_{21}^* \end{bmatrix} \tag{19}$$

Equations (14) and (18) can be combined to yield the matrix form

$$\mathbf{r} = \begin{bmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix} \begin{bmatrix} \mathbf{c} \\ \mathbf{s} \end{bmatrix} + \begin{bmatrix} \mathbf{n}_1 \\ \mathbf{n}_2 \end{bmatrix}. \tag{20}$$

<u>Zero-Forcing IC and ML Decoding Scheme:</u> In this case, a matrix $\mathbf{W}$ is chosen such that

$$\mathbf{W} \cdot \begin{bmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{r}}_1 \\ \tilde{\mathbf{r}}_2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{H}} & \mathbf{0} \\ \mathbf{0} & \tilde{\mathbf{G}} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{c} \\ \mathbf{s} \end{bmatrix} + \begin{bmatrix} \tilde{\mathbf{n}}_1 \\ \tilde{\mathbf{n}}_2 \end{bmatrix} \tag{21}$$

We can find an appropriate matrix $\mathbf{W}$ by realizing that

$$\begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix}^{-1} = \begin{bmatrix} \mathbf{A}_1^{-1} & \mathbf{0} \\ \mathbf{0} & \mathbf{A}_2^{-1} \end{bmatrix} \begin{bmatrix} \mathbf{I} & -\mathbf{G}_1\mathbf{G}_2^{-1} \\ -\mathbf{H}_2\mathbf{H}_1^{-1} & \mathbf{I} \end{bmatrix} \tag{22}$$

where

$$\mathbf{A}_1 = \mathbf{H}_1 - \mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{H}_2 \quad \text{and} \quad \mathbf{A}_2 = \mathbf{G}_2 - \mathbf{H}_2\mathbf{H}_1^{-1}\mathbf{G}_1 \tag{23}$$

Hence, if we select $\mathbf{W}$ as

$$\mathbf{W} = \begin{bmatrix} \mathbf{I} & -\mathbf{G}_1\mathbf{G}_2^{-1} \\ -\mathbf{H}_2\mathbf{H}_1^{-1} & \mathbf{I} \end{bmatrix} \tag{24}$$

we will have

$$\mathbf{W}\cdot\begin{bmatrix}\mathbf{r}_1\\ \mathbf{r}_2\end{bmatrix}=\begin{bmatrix}\widetilde{\mathbf{r}}_1\\ \widetilde{\mathbf{r}}_2\end{bmatrix}=\begin{bmatrix}\widetilde{\mathbf{H}} & \mathbf{0}\\ \mathbf{0} & \widetilde{\mathbf{G}}\end{bmatrix}\cdot\begin{bmatrix}\mathbf{c}\\ \mathbf{s}\end{bmatrix}+\begin{bmatrix}\widetilde{\mathbf{n}}_1\\ \widetilde{\mathbf{n}}_2\end{bmatrix} \tag{25}$$

where

$$\begin{aligned}\widetilde{\mathbf{H}} &= \mathbf{H}_1-\mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{H}_2\\ \widetilde{\mathbf{G}} &= \mathbf{G}_2-\mathbf{H}_2\mathbf{H}_1^{-1}\mathbf{G}_1\\ \widetilde{\mathbf{n}}_1 &= \mathbf{n}_1-\mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{n}_2\\ \widetilde{\mathbf{n}}_2 &= \mathbf{n}_2-\mathbf{H}_2\mathbf{H}_1^{-1}\mathbf{n}_1\end{aligned} \tag{26}$$

From equation (25) it can be easily observed that the modified received signal vector $\widetilde{\mathbf{r}}_1$ contains signals only from transmitter 10 (*i.e.* signals from transmitter 30 have been *canceled* or removed) and, correspondingly, the modified received signal vector $\widetilde{\mathbf{r}}_2$ contains signals only from transmitter 30 (*i.e.* signals from transmitter 10 have been *canceled* or removed). A number of other attributes can be shown from the above to be true.

1) the modified noise vector $\widetilde{\mathbf{n}}_1$ is a zero mean complex Gaussian random vector with covariance

$$\mathbf{R}_{n1}=N_o\cdot\left(1+\frac{D_{g1}}{D_{g2}}\right)\cdot\mathbf{I} \tag{27}$$

where $D_{g1}=|g_{11}|^2+|g_{12}|^2$ and $D_{g2}=|g_{21}|^2+|g_{22}|^2$. Hence, the modified noise vector $\widetilde{\mathbf{n}}_1$ is also white.

2) the modified noise vector $\widetilde{\mathbf{n}}_2$ is also a zero mean Gaussian random vector with covriance

$$\mathbf{R}_{n2}=N_o\cdot\left(1+\frac{D_{h2}}{D_{h1}}\right)\cdot\mathbf{I} \tag{28}$$

where $D_{h1}=|h_{11}|^2+|h_{12}|^2$ and $D_{h2}=|h_{21}|^2+|h_{22}|^2$, and hence it is also white.

3) The matrices $\widetilde{\mathbf{H}}$ and $\widetilde{\mathbf{G}}$ the form

$$\widetilde{\mathbf{H}}=\begin{bmatrix}\widetilde{h}_1 & \widetilde{h}_2\\ \widetilde{h}_2^* & -\widetilde{h}_1^*\end{bmatrix},\ \text{and}\ \widetilde{\mathbf{G}}=\begin{bmatrix}\widetilde{g}_1 & \widetilde{g}_2\\ \widetilde{g}_2^* & -\widetilde{g}_1^*\end{bmatrix}. \tag{29}$$

4) Conditioned on $\mathbf{G}_1$ and $\mathbf{G}_2$, the random variables $\tilde{h}_1$ and $\tilde{h}_2$ are both zero mean complex Gaussian random variables with variance $\sigma_h^2 = 1 + D_{g1}/D_{g2}$.

5) Conditioned on $\mathbf{H}_1$ and $\mathbf{H}_2$, the random variables $\tilde{g}_1$ and $\tilde{g}_2$ are both zero mean complex Gaussian random variables with variance $\sigma_g^2 = 1 + D_{h2}/D_{h1}$.

6) The modified channel matrices $\tilde{\mathbf{H}}$ and $\tilde{\mathbf{G}}$ have a structure similar to that in equation (5), i.e. the modified channel matrices $\tilde{\mathbf{H}}$ and $\tilde{\mathbf{G}}$ are orthogonal matrices.

Considering the modified received signal vector $\tilde{\mathbf{r}}_1$, which contains signals only from transmitter 10, i.e.,

$$\tilde{\mathbf{r}}_1 = \tilde{\mathbf{H}} \cdot \mathbf{c} + \tilde{\mathbf{n}}_1, \tag{30}$$

it is noted that this expression resembles the expression in equation (4). Hence, in accordance with the principles disclosed therein, the optimum ML decoder for the symbols transmitted by terminal unit 10 evaluates an equation that is similar to the expression of equation (6), and is given by

$$\hat{\mathbf{c}} = \arg\min_{\hat{c} \in C} \left\| \tilde{\mathbf{r}}_1 - \tilde{\mathbf{H}} \cdot \hat{\mathbf{c}} \right\|^2. \tag{31}$$

The corresponding decoder uncertainty is given by

$$\Delta_c = \left\| \tilde{\mathbf{r}}_1 - \tilde{\mathbf{H}} \cdot \hat{\mathbf{c}} \right\|^2. \tag{32}$$

Moreover, since the channel Matrix $\tilde{\mathbf{H}}$ is orthogonal, the ML decoder will also decompose into two separate rules for c_1 and c_2.

In a similar fashion, considering the modified received signal vector $\tilde{\mathbf{r}}_2$, which contains signals only from transmitter 10, i.e.,

$$\tilde{\mathbf{r}}_2 = \tilde{\mathbf{H}} \cdot \mathbf{c} + \tilde{\mathbf{n}}_2, \tag{33}$$

it is noted that this expression resembles the expression in equation (4). Hence, the optimum ML decoder for the symbols transmitted by terminal unit 30 evaluates an equation that is similar to the expression of equation (6), and is given by

$$\hat{\mathbf{s}} = \arg\min_{\hat{s} \in S} \left\| \tilde{\mathbf{r}}_2 - \tilde{\mathbf{G}} \cdot \hat{\mathbf{s}} \right\|^2. \tag{34}$$

The corresponding decoder uncertainty is given by

$$\Delta_s = \left\| \tilde{\mathbf{r}}_2 - \tilde{\mathbf{G}} \cdot \hat{\mathbf{s}} \right\|^2. \qquad (35)$$

Moreover, since the channel Matrix $\tilde{\mathbf{G}}$ is also orthogonal, the ML decoder will also decompose into two separate rules for s_1 and s_2.

The above-disclosed technique can be easily implemented within a detector 25 that comprises a stored program general purpose processor. Specifically, a subroutine $(\hat{\mathbf{c}}, \Delta)$=ZF.DECODE($\mathbf{r}_1$, $\mathbf{r}_2$, $\mathbf{H}_1$, $\mathbf{H}_2$, $\mathbf{G}_1$, $\mathbf{G}_2$) can be installed which returns the values $\hat{\mathbf{c}}, \Delta$ in response to submitted inputs $\mathbf{r}_1, \mathbf{r}_2, \mathbf{H}_1, \mathbf{H}_2, \mathbf{G}_1$, and $\mathbf{G}_2$, as shown below:

$$(\hat{\mathbf{c}}, \Delta_c) = \text{ZF.DECODE}(\mathbf{r}_1, \mathbf{r}_2, \mathbf{H}_1, \mathbf{H}_2, \mathbf{G}_1, \mathbf{G}_2)$$

$$\{$$

$$\tilde{\mathbf{r}} = \mathbf{r}_1 - \mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{r}_2$$

$$\tilde{\mathbf{H}} = \mathbf{H}_1 - \mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{H}_2$$

$$\hat{\mathbf{c}} = \arg\min_{\mathbf{c} \in \mathbf{C}} \left\| \tilde{\mathbf{r}} - \tilde{\mathbf{H}} \cdot \mathbf{c} \right\|^2$$

$$\Delta_c = \left\| \tilde{\mathbf{r}} - \tilde{\mathbf{H}} \cdot \hat{\mathbf{c}} \right\|^2$$

$$\}$$

With such a subroutine, both $\hat{\mathbf{s}}$ and $\hat{\mathbf{c}}$ can be estimated. as follows:

$$(\hat{\mathbf{c}}, \Delta) = \text{ZF.DECODE}(\mathbf{r}_1, \mathbf{r}_2, \mathbf{H}_1, \mathbf{H}_2, \mathbf{G}_1, \mathbf{G}_2) \qquad (36)$$

$$(\hat{\mathbf{s}}, \Delta) = \text{ZF.DECODE}(\mathbf{r}_2, \mathbf{r}_1, \mathbf{G}_2, \mathbf{G}_1, \mathbf{H}_2, \mathbf{H}_1). \qquad (37)$$

It may be noted that knowledge of the noise power N_0 is not required. Simulation results reveal that the performance of the FIG. 1 system which employs the principles disclosed herein and the ZF.DECODE subroutine is equivalent to that when only one terminal unit exists and the base station uses a single receive antenna. That is, the reception is equivalent to the performance of two branch MRRC diversity receiver. Advantageously, however, the disclosed technique is able to support two *co-channel* terminal units.

The discussion above centered on describing the technique for canceling out the signal of transmitter 10 when detecting the signal of

transmitter 30, and for canceling out the signal of transmitter 30 when detecting the signal of transmitter 10. Effectively, detector 25 of receiver 20 can either comprise two processors, with one making the subroutine call of equation (31) and the other making the subroutine call of equation (32). Alternatively, the signals can be stored within detector 25 and the subroutine calls of equations 31 and 32 can be made seriatim.

Minimum Mean-Squared Error IC and ML Decoding Scheme: The above-disclosed approach for canceling the contribution of an interfering terminal unit is known as the zero-forcing (ZF) as a minimum mean-squared error technique (MMSE).

Recalling equation (20), the vector $\mathbf{r}$ can also be written as

$$\mathbf{r} = \mathbf{H} \cdot \tilde{\mathbf{c}} + \mathbf{n} \tag{38}$$

where $\tilde{\mathbf{c}} = \left[\tilde{\mathbf{c}}^T \quad \tilde{\mathbf{s}}^T\right]^T$, $\mathbf{r} = \left[\tilde{\mathbf{r}}_1^T \quad \tilde{\mathbf{r}}_2^T\right]^T = \left[r_{11} \quad r_{21}^* \quad r_{12} \quad r_{22}^*\right]^T$, and

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix} \tag{39}$$

To simplify notations, the vector $\mathbf{r}$ is also defined as $\mathbf{r} = \left[r_1 \quad r_2 \quad r_3 \quad r_4\right]^T$.

When seeking to detect and decode signals $\{c_1, c_2\}$ by minimizing a mean-squared error criterion, the goal is find a linear combination of the received signals such that the mean-squared error in detecting the signals $\{c_1, c_2\}$ is minimized. In general terms, this can be expressed by an error cost function that is to be minimized, such as the function

$$J(\alpha, \beta) = \left\| \sum_{i=1}^{4} \alpha_i r_i - (\beta_1 c_1 + \beta_2 c_2) \right\|^2 = \left\| \alpha \cdot \mathbf{r} - \beta \cdot \mathbf{c} \right\|^2 \tag{40}$$

One may note that a minimum is certainly reached when both α and β are equal to 0, but that, of course, is not desired. Therefore, either β_1 or β_2 is set to 1. When β_2 is set to 1, we get the following minimization criterion from equation (40)

$$J_1(\alpha_1, \beta_1) = \left\| \sum_{i=1}^{5} \alpha_{1i} r_{1i} - c_2 \right\|^2 = \left\| \tilde{\alpha}_1 \tilde{\mathbf{r}}_1 - c_2 \right\|^2 \tag{41}$$

where $\tilde{\alpha}_1 = [\alpha_{11}, \alpha_{12}, \alpha_{13}, \alpha_{14}, -\beta_1] = [\alpha_1 \quad -\beta_1]$ and $\tilde{\mathbf{r}}_1 = \left[\mathbf{r}^T \quad c_1\right]^T$. From this it

can be seen that

$$\tilde{\mathbf{r}}_1 = \begin{bmatrix} \mathbf{H} & \mathbf{0}^T \\ \mathbf{0} & 1 \end{bmatrix} \begin{bmatrix} \tilde{\mathbf{c}} \\ c_1 \end{bmatrix} + \begin{bmatrix} \mathbf{n} \\ 0 \end{bmatrix} = \mathbf{R} \cdot \mathbf{d} + \eta \qquad (42)$$

where $\mathbf{0} = [0 \quad 0 \quad 0 \quad 0]$.

What is needed is to select $\tilde{\alpha}_1$ so that the expected value of the expression in equation (41) is minimized. That is, select $\tilde{\alpha}_1$ to minimize

$$E\{J_1(\tilde{\alpha}_1)\} = E\{(\tilde{\alpha}_1\tilde{\mathbf{r}}_1 - c_2)(\tilde{\alpha}_1\tilde{\mathbf{r}}_1 - c_2)^*\}. \qquad (43)$$

Taking the partial derivative with respect to $\tilde{\alpha}_1$ and setting it to zero, what results is

$$\begin{bmatrix} \mathbf{M} & \mathbf{h}_1 \\ \mathbf{h}_1^* & 1 \end{bmatrix} \begin{bmatrix} \alpha_1^* \\ -\beta_1 \end{bmatrix} = \begin{bmatrix} \mathbf{h}_2 \\ 0 \end{bmatrix} \qquad (44)$$

where $\mathbf{M} = \mathbf{H}\mathbf{H}^* + \frac{1}{\Gamma}\mathbf{I}$, Γ is the signal to noise ratio, $\mathbf{I}$ is the 4 by 4 identity matrix, $\mathbf{h}_1$ is the first column of $\mathbf{H}$, and $\mathbf{h}_2$ is the second column of $\mathbf{H}$. It follows that

$$\alpha_1^* = (\mathbf{M} - \mathbf{h}_1\mathbf{h}_1^*)^{-1}\mathbf{h}_2 \quad \text{and} \quad \beta_1^* = \mathbf{h}_1^*(\mathbf{M} - \mathbf{h}_1\mathbf{h}_1^*)^{-1}\mathbf{h}_2. \qquad (45)$$

It can be shown that

$$(\mathbf{M} - \mathbf{h}_1\mathbf{h}_1^*)^{-1} = \mathbf{M}^{-1} - \frac{\mathbf{M}^{-1}\mathbf{h}_1\mathbf{h}_1^*\mathbf{M}^{-1}}{1 - \mathbf{h}_1^*\mathbf{M}^{-1}\mathbf{h}_1}, \qquad (46)$$

which yields

$$\beta_1^* = \frac{\mathbf{h}_1^*\mathbf{M}^{-1}\mathbf{h}_2}{1 - \mathbf{h}_1^*\mathbf{M}^{-1}\mathbf{h}_1} \qquad (47)$$

From the structure of the matrix $\mathbf{H}$ we can easily verify that $\mathbf{h}_1$ and $\mathbf{h}_2$ are orthogonal. Using this fact and the structure of the matrix $\mathbf{M}$, it can be shown that

$$\beta_1 = 0 \qquad (48)$$

$$\alpha_1^* = \mathbf{M}^{-1}\mathbf{h}_2 \qquad (49)$$

Hence, the MMSE IC solution given in equations (45) and (46) will minimize the mean-squared error in c_2 without any regard to c_1. Considering the alternative cost function when β_1 is set to 1, a similar analysis leads to the

conclusion that

$$\beta_2 = 0 \tag{47}$$

$$\alpha_2^* = \mathbf{M}^{-1}\mathbf{h}_1 \tag{48}$$

The value of Γ and the values of h_{ij} and g_{ij}, and consequently, the values of **H** and **M** are obtained from a training sequence in a conventional manner by elements 23 and 24. Since, as indicated earlier, this is quite conventional and does not form a part of this invention, for sake of conciseness additional details are not presented. In this case, the MMSE IC solution given in equations (45) and (46) will minimize the mean-squared error in c_1 without any regard to c_2. Therefore, from equations (45)-(48), we can easily see that the MMSE interference canceller for signals from terminal unit 10 will consist of two different sets of weights α_1 and α_2 for c_2 and c_1, respectively. The weights for decoding signals from terminal 30 can be obtained in a similar fashion, as expected. Thus, the decoding of signals from terminal units 10 and 30 can be performed with a single subroutine MMSE.DECODE in decoder 25 as follows:

$$(\mathbf{c}, \Delta_c) = \text{MMSE.DECODE}(\mathbf{r}_1, \mathbf{r}_2, \mathbf{H}_1, \mathbf{H}_2, \mathbf{G}_1, \mathbf{G}_2, \Gamma)$$

{

$$\tilde{\mathbf{r}} = \begin{bmatrix} \mathbf{r}_1^T & \mathbf{r}_2^T \end{bmatrix}^T$$

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix}$$

$$\mathbf{M} = \mathbf{H}\mathbf{H}^* + \frac{1}{\Gamma}\mathbf{I}$$

$$\mathbf{h}_1 = \begin{bmatrix} \mathbf{h}_{11}^T & \mathbf{h}_{21}^T \end{bmatrix}^T = \text{ first column of } \mathbf{H}$$

$$\mathbf{h}_2 = \begin{bmatrix} \mathbf{h}_{12}^T & \mathbf{h}_{22}^T \end{bmatrix}^T = \text{ second column of } \mathbf{H}$$

$$\alpha_1^* = \mathbf{M}^{-1}\mathbf{h}_1 \text{ , } \alpha_2^* = \mathbf{M}^{-1}\mathbf{h}_2$$

$$c_1 = \arg\min_{\hat{c}_1 \in C} \left\|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_1\right\|^2, \quad c_2 = \arg\min_{\hat{c}_2 \in C} \left\|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_2\right\|^2$$

$$\Delta_c = \left\|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_1\right\|^2 + \left\|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_2\right\|^2$$

}

With such a subroutine, both $\hat{s}$ and $\hat{c}$ can be estimated, as follows:

$$(\hat{c},\Delta)=\text{MMSE.DECODE}(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2,\Gamma) \tag{49}$$

$$(\hat{s},\Delta)=\text{MMSE.DECODE}(\mathbf{r}_2,\mathbf{r}_1,\mathbf{G}_1,\mathbf{G}_2,\mathbf{H}_1,\mathbf{H}_2,\Gamma) \tag{50}$$

Similar to the zero-forcing case, simulation results reveal that the performance of the disclosed technique MMSE.DECODE is equivalent to that when only one terminal unit exists and the base station uses a single receive antenna which is equivalent to the performance of two branch MRRC diversity. However, this technique is also able to support two *co-channel* terminal units. In addition, when the SIR (signal-to-interference ratio, which is a ratio between the desired terminal power to the interfering terminal power) increases, the MMSE approach will have a better performance as compared to the ZF case (the ZF performance is almost the same as the performance of the MMSE approach at 0 dB SIR).

Two-Step Interference Cancellation: Actually, additional improvement can be realized by employing a two-step interference cancellation approach using either the zero-forcing or the MMSE interference cancellation techniques disclosed above. Below, we will describe this approach based on the MMSE technique. However, as one might expect there a similar approach based on the zero-forcing technique. In this two-step approach, the receiver decodes signals from both terminals using the subroutine MMSE.DECODE disclosed above. Assuming that symbols from the terminal unit 10, $\hat{c}_o$, have been decoded correctly, the receiver can, then, *perfectly* cancel the contribution of the terminal unit 10 in the received signal vectors $\mathbf{r}_1$ and $\mathbf{r}_2$. The receiver then uses $\mathbf{x}_1$ and $\mathbf{x}_2$, the received signal vectors after canceling signals from terminal unit 10, to *re-decode* symbols from terminal unit 30 $\hat{s}_o$ using the optimum ML decoding rule in equation (10). Assuming that the symbols from terminal unit 10 has been decoded correctly, we can easily see that, the performance for terminal

unit 30 will be equivalent to that with 2 transmit and 2 receive antennas (which is equivalent to 4 branch MRC diversity). If we let $\Delta_o = \Delta_{c_o} + \Delta_{s_o}$ denotes the overall uncertainty for $\hat{\mathbf{c}}_o$ and $\hat{\mathbf{s}}_o$. The receiver then repeats the above step assuming that symbols from terminal unit 30 $\hat{\mathbf{s}}_1$ have been decoded correctly using the MMSE.DECODE subroutine. As before, the receiver cancels the contribution of terminal unit 30 in the received signal vectors $\mathbf{r}_1$ and uses $\mathbf{y}_1$ and $\mathbf{y}_2$, the received signal vectors after cancelling signals from terminal unit 30, *to re-decode* symbols from terminal unit 10 $\hat{\mathbf{c}}_1$ using the optimum ML decoding rule in equation (10). As before, assuming that symbols from terminal unit 30, the performance for terminal unit 10 will be equivalent to that with 2 transmit and 2 receive antennas. Similarly, let $\Delta_1 = \Delta_{c_1} + \Delta_{s_1}$ denotes the overall uncertainty for $\hat{\mathbf{c}}_1$ and $\hat{\mathbf{s}}_1$. The receiver then compares the overall uncertainty and chooses the pair $(\hat{\mathbf{c}}_o, \hat{\mathbf{s}}_o)$ if $\Delta_o < \Delta_1$ and $(\hat{\mathbf{c}}_1, \hat{\mathbf{s}}_1)$ otherwise. The two-step interference cancellation approach based on the MMSE technqiue disclosed above is presented below in pseudo-code subroutine II.MMSE.DECODE. As we mentioned earlier, the techniques can be also used with the zero forcing approach. Below, we also present the pseudo code subroutine II.ZF.DECODE for the two-step interference cancellation based on the zero-forcing approach.

$$(\hat{\mathbf{c}},\hat{\mathbf{s}}) = \text{II. MMSE.DECODE}(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2,\Gamma)$$

{

$$(\hat{\mathbf{c}}_0,\Delta_{c,o}) = \text{MMSE.DECODE}(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2,\Gamma)$$

$$\mathbf{x}_1 = \mathbf{r}_1 - \mathbf{H}_1\cdot\hat{\mathbf{c}}_o \quad , \quad \mathbf{x}_2 = \mathbf{r}_2 - \mathbf{H}_2\cdot\hat{\mathbf{c}}_o$$

$$\hat{\mathbf{s}}_o = \arg\min_{s\in\mathbf{S}} \left(\|\mathbf{x}_1 - \mathbf{G}_1\cdot\mathbf{s}\|^2 + \|\mathbf{x}_2 - \mathbf{G}_2\cdot\mathbf{s}\|^2\right)$$

$$\Delta_{s,o} = \|\mathbf{x}_1 - \mathbf{G}_1\cdot\mathbf{s}\|^2 + \|\mathbf{x}_2 - \mathbf{G}_2\cdot\mathbf{s}\|^2$$

$$(\hat{\mathbf{s}}_1,\Delta_{s,1}) = \text{MMSE.DECODE}(\mathbf{r}_2,\mathbf{r}_1,\mathbf{G}_2,\mathbf{G}_1,\mathbf{H}_1,\mathbf{H}_2,\Gamma)$$

$$\mathbf{y}_1 = \mathbf{r}_1 - \mathbf{G}_1\cdot\hat{\mathbf{s}}_1 \quad , \quad \hat{\mathbf{y}}_2 = \mathbf{r}_2 - \mathbf{G}_2\cdot\hat{\mathbf{s}}_1$$

$$\hat{\mathbf{c}}_1 = \arg\min_{c\in\mathbf{C}} \left(\|\mathbf{y}_1 - \mathbf{H}_1\cdot\mathbf{c}\|^2 + \|\mathbf{y}_2 - \mathbf{H}_2\cdot\mathbf{c}\|^2\right)$$

$$\Delta_{c,1} = \|\mathbf{y}_1 - \mathbf{H}_1\cdot\mathbf{c}\|^2 + \|\mathbf{y}_2 - \mathbf{H}_2\cdot\mathbf{c}\|^2$$

If $(\Delta_{c,o} + \Delta_{s,o}) < (\Delta_{c,1} + \Delta_{s,1}) \Rightarrow (\hat{\mathbf{c}},\hat{\mathbf{s}}) = (\hat{\mathbf{c}}_o,\hat{\mathbf{s}}_o)$

Else $(\hat{\mathbf{c}},\hat{\mathbf{s}}) = (\hat{\mathbf{c}}_1,\hat{\mathbf{s}}_1)$

}

$$(\hat{\mathbf{c}},\hat{\mathbf{s}}) = \text{II. ZF.DECODE}(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2)$$

$$(\hat{\mathbf{c}}_0,\Delta_{c,o}) = \text{ZF.DECODE}(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2)$$

$$\mathbf{x}_1 = \mathbf{r}_1 - \mathbf{H}_1\cdot\hat{\mathbf{c}}_o \quad , \quad \mathbf{x}_2 = \mathbf{r}_2 - \mathbf{H}_2\cdot\hat{\mathbf{c}}_o$$

$$\hat{\mathbf{s}}_o = \arg\min_{s\in\mathbf{S}} \left(\|\mathbf{x}_1 - \mathbf{G}_1\cdot\mathbf{s}\|^2 + \|\mathbf{x}_2 - \mathbf{G}_2\cdot\mathbf{s}\|^2\right)$$

$$\Delta_{s,o} = \|\mathbf{x}_1 - \mathbf{G}_1\cdot\mathbf{s}\|^2 + \|\mathbf{x}_2 - \mathbf{G}_2\cdot\mathbf{s}\|^2$$

$$(\hat{\mathbf{s}}_1,\Delta_{s,1}) = \text{ZF.DECODE}(\mathbf{r}_2,\mathbf{r}_1,\mathbf{G}_2,\mathbf{G}_1,\mathbf{H}_2,\mathbf{H}_1)$$

$$\mathbf{y}_1 = \mathbf{r}_1 - \mathbf{G}_1\cdot\hat{\mathbf{s}}_1 \quad , \quad \hat{\mathbf{y}}_2 = \mathbf{r}_2 - \mathbf{G}_2\cdot\hat{\mathbf{s}}_1$$

$$\hat{\mathbf{c}}_1 = \arg\min_{c\in\mathbf{C}} \left(\|\mathbf{y}_1 - \mathbf{H}_1\cdot\mathbf{c}\|^2 + \|\mathbf{y}_2 - \mathbf{H}_2\cdot\mathbf{c}\|^2\right)$$

$$\Delta_{c,1} = \|\mathbf{y}_1 - \mathbf{H}_1\cdot\mathbf{c}\|^2 + \|\mathbf{y}_2 - \mathbf{H}_2\cdot\mathbf{c}\|^2$$

If $(\Delta_{c,o} + \Delta_{s,o}) < (\Delta_{c,1} + \Delta_{s,1}) \Rightarrow (\hat{\mathbf{c}},\hat{\mathbf{s}}) = (\hat{\mathbf{c}}_o,\hat{\mathbf{s}}_o)$

Else $(\hat{\mathbf{c}},\hat{\mathbf{s}}) = (\hat{\mathbf{c}}_1,\hat{\mathbf{s}}_1)$

Interference Cancellation and ML Decoding: GENERAL CASE

In the above basic case, we focused on the basic case where we assumed two co-channel terminals (K=2) each uses two transmit antennas (N=2). Both terminals communicate with a base station that is equipped with two transmit antennas (M=2). In this section we will consider the more general case of $K \geq 2$ co-channel terminals each is equipped with $N \geq 2$ transmitting antennas, both terminals communicate with a base that has receive $M \geq K$ antennas. We will develop similar interference cancellation and ML decoding scheme for this case.

In a paper submitted to IEEE Transactions on Information Theory, Vahid Tarokh et al. extended the above space-time block coding scheme to the case when more than two antennas are used for transmission ($N \geq 2$). There, a technique for constructing space-time block codes (with similar properties to the simple scheme described above) was developed. It was also shown that for real constellations space-time block codes with transmission rate 1 can be constructed. ***However, for a general complex constellation the rate of transmission for these codes will be less than 1.***

In general, let us assume that the input information symbols to the transmitter are grouped into groups of Q symbols $c_1, c_2, \cdots, c_Q$. A space-time block code in this case will map the symbols $c_1, c_2, \cdots, c_Q$ into an $N \times L$ array $\mathbf{C}$ whose entries are made $\pm c_1, \pm c_2, \cdots, \pm c_Q$ and $\pm c_1^*, \pm c_2^*, \cdots, \pm c_Q^*$. At time t, where $1 \leq t \leq L$, the t-th column of $\mathbf{C}$ is transmitted from the N antennas. In this case, the transmission rate of such code will be Q/L. In general, for a rate Q/L space-time block code (as constructed by V. Tarokh et al.) designed for N transmit antenna, let $r_1, r_2, \cdots, r_L$ be the received signals at time $t = 1, 2, \cdots, L$. As before, we define the received signal vector as

$$\mathbf{r} = \begin{bmatrix} r_1 & r_2 & \cdots & r_{L/2} & r_{L/2+1}^* & r_{L/2+2}^* & \cdots & r_L^* \end{bmatrix}^T \tag{51}$$

where the $L \times 1$ vector $\mathbf{r}$ can be written as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \eta \tag{52}$$

and $\mathbf{H}$ is the $L \times Q$ channel matrix whose entries are

from $\pm h_1, \pm h_2, \cdots, \pm h_N, \pm h_1^*, \pm h_2^*, \cdots, \pm h_N^*$, and it is an orthogonal matrix, $\mathbf{c} = \begin{bmatrix} c_1 & c_2 & \cdots & c_Q \end{bmatrix}^T$, and η is an $L \times 1$ zero-mean complex Gaussian random vector with covariance $N_o \cdot \mathbf{I}$ which models the noise. The ML decoder in this case is similar to that in equation (6), that is

$$\hat{\mathbf{c}} = \arg \min_{\hat{\mathbf{c}} \in \mathbf{C}} \left\| \mathbf{r} - \mathbf{H} \cdot \hat{\mathbf{c}} \right\|^2 \tag{53}$$

and the *uncertainty,* Δ_c, of the decoded symbols $\hat{\mathbf{c}}$ is given by

$$\Delta_c = \left\| \mathbf{r} - \mathbf{H} \cdot \hat{\mathbf{c}} \right\|^2 . \tag{54}$$

As before, since the channel matrix $\mathbf{H}$ is orthogonal, the decoding rule in (53) decomposes into Q separate decoding rules for $c_1, c_2, \cdots, c_Q$. For example, assuming that the terminal unit uses 4 transmit antenna, a rate 4/8 (i.e. it is a rate ½) space-time block code is given by

$$\begin{bmatrix} c_1 \\ c_2 \\ c_3 \\ c_4 \end{bmatrix} \rightarrow \begin{bmatrix} c_1 & -c_2 & -c_3 & -c_4 & c_1^* & -c_2^* & -c_3^* & -c_4^* \\ c_2 & c_1 & c_4 & -c_3 & c_2^* & c_1^* & c_4^* & -c_3^* \\ c_3 & -c_4 & c_1 & c_2 & c_3^* & -c_4^* & c_1^* & c_2^* \\ c_4 & c_3 & -c_2 & c_1 & c_4^* & c_3^* & -c_2^* & c_1^* \end{bmatrix}$$

In this case, at time t=1 c_1, c_2, c_3, c_4 are transmitted from antenna 1 through 4, respectively. At time t=2, $-c_2, c_1, -c_4, c_3$ are transmitted from antenna 1 through 4, respectively, and so on. For this example, let $r_1, r_2, \cdots, r_8$ be the received signals at time $t = 1,2,\cdots,8$. Define the received signal vector $\mathbf{r} = \begin{bmatrix} r_1 & r_2 & r_3 & r_4 & r_5^* & r_6^* & r_7^* & r_8^* \end{bmatrix}^T$. In this case, we can write the received signal vector $\mathbf{r}$ can be written as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \eta \tag{55}$$

where η is the 8×1 AWGN noise vector and $\mathbf{H}$ is the 8×4 channel matrix given by:

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2 & -h_1 & h_4 & -h_3 \\ h_3 & -h_4 & -h_1 & h_2 \\ h_4 & h_3 & -h_2 & -h_1 \\ h_1^* & h_2^* & h_3^* & h_4^* \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & -h_4^* & -h_1^* & h_2^* \\ h_4^* & h_3^* & -h_2^* & -h_1^* \end{bmatrix} \qquad (56)$$

We can immediately notice that the matrix $\mathbf{H}$ is orthogonal, that is $\mathbf{H}^*\mathbf{H} = D_h \cdot \mathbf{I}$ where $D_h = \sum_{i=1}^{4} |h_i|^2$ and $\mathbf{I}$ is a 4×4 identity matrix.

Let us now assume a multi-user environment with K co-channel synchronized terminals. Each terminal uses a rate Q/L space-time block code with N transmit antenna (as constructed by V. Tarokh et al). The base station uses $M \geq K$ antennas for reception. We can write the received signal vector at the m-th receive antenna as

$$\mathbf{r}_m = \sum_{k=1}^{K} \mathbf{H}_{km} \cdot \mathbf{c}_k + \eta_m, \qquad m = 1,2,\cdots,M \qquad (57)$$

where $\mathbf{H}_{km}$ is the $L \times Q$ k-th user channel matrix to antenna m, $\mathbf{c}_k = \begin{bmatrix} c_{k1} & c_{k2} & \cdots & c_{kQ} \end{bmatrix}^T$ is the $Q \times 1$ information symbols vector for k-th user, and η_m is the $L \times 1$ noise vector. The entries of the k-th user channel matrix $\mathbf{H}_{km}$ are from $\pm h_{k,m,1}, \pm h_{k,m,2}, \cdots, \pm h_{k,m,N}$ and $\pm h_{k,m,1}^*, \pm h_{k,m,2}^*, \cdots, \pm h_{k,m,N}^*$, where $h_{k,m,n}$ is the complex channel gain between transmit antenna n of the k-th user and receive antenna m. As we stated earlier, the matrix $\mathbf{H}_{km}$ is orthogonal.

Zero-Forcing IC and ML Decoding: Without loss of generality, let us assume that we are interested in suppressing signals from co-channel terminals $2,3,\cdots,K$ while decoding signals from the first terminal. This can be done in a successive manner as follows.

First, let us define $\mathbf{r}_m^{(0)} = \mathbf{r}_m$. Let us assume that we start by canceling out the contributions of the K-th terminal. We can use the M-th antenna received signal vector $\mathbf{r}_M$ to cancel out the contribution of the K-th terminal in the remaining M-1 received signal vectors by forming the modified received signal vectors $\mathbf{r}_m^{(1)}, m = 1,\cdots,M-1$ as follows:

$$\mathbf{r}_m^{(1)} = \mathbf{r}_m^{(0)} - \mathbf{H}_{Km}\mathbf{H}_{KM}^{+}\mathbf{r}_M^{(0)} \qquad m = 1,2,\cdots M-1 \tag{58}$$

where $\mathbf{H}_{km}^{+}$ is the generalized inverse of the channel matrix $\mathbf{H}_{km}$ and is given by

$$\mathbf{H}_{km}^{+} = \left(\mathbf{H}_{km}^{*}\mathbf{H}_{km}\right)^{-1}\mathbf{H}_{km}^{*} \tag{59}$$

We can easily verify that $\mathbf{H}_{km}^{+}\mathbf{H}_{km} = \mathbf{I}$, where $\mathbf{I}$ is the $Q\times Q$ identity matrix. We can easily verify that the modified received signal vectors $\mathbf{r}_m^{(1)}, m = 1,\cdots,M-1$, do not contain any signal contribution due to the K-th user. Moreover, we can easily verify that $\mathbf{r}_m^{(1)}$ can be written as

$$\mathbf{r}_m^{(1)} = \sum_{k=1}^{K-1}\mathbf{H}_{km}^{(1)}\cdot\mathbf{c}_k + \eta_m^{(1)}, \quad m = 1,2,\cdots,M-1 \tag{60}$$

where $\mathbf{H}_{km}^{(1)}$ and $\eta_m^{(1)}$ are given by

$$\mathbf{H}_{km}^{(1)} = \mathbf{H}_{km}^{(0)} - \mathbf{H}_{Km}^{(0)}\left(\mathbf{H}_{KM}^{(0)}\right)^{+}\mathbf{H}_{kM}^{(0)}, \quad m = 1,2,\cdots,M-1 \tag{61}$$

$$\eta_m^{(1)} = \eta_m^{(0)} - \mathbf{H}_{Km}^{(0)}\left(\mathbf{H}_{KM}^{(0)}\right)^{+}\eta_M^{(0)}, \quad m = 1,2,\cdots,M-1 \tag{62}$$

Moreover, it can be shown that for codes constructed by V. Tarokh et al, the modified channel matrix $\mathbf{H}_{km}^{(1)}$ ***will have exactly the same structure as that of*** $\mathbf{H}_{km}$. That is, the entries of the k-th user modified channel matrix $\mathbf{H}_{km}^{(1)}$ are from $\pm h_{k,m,1}^{(1)}, \pm h_{k,m,2}^{(1)}, \cdots, \pm h_{k,m,N}^{(1)}$ and $\pm h_{k,m,1}^{(1)*}, \pm h_{k,m,2}^{(1)*}, \cdots, \pm h_{k,m,N}^{(1)*}$, where $h_{k,m,n}^{(1)}$ is the modified complex channel gain between transmit antenna n of the k-th user and receive antenna m, m=1,...,M-1. Hence, the modified channel matrix $\mathbf{H}_{km}^{(1)}$ will be orthogonal as well.

It can then be observed that the expression for the M-1 modified received signal vector $\mathbf{r}_m^{(1)}$ in equation (60) is the same as that in equation (57) except that we now have one less interfering terminal. In a similar fashion, we can cancel out the contributions of terminal K-1 and obtain M-2

modified received signal vectors $\mathbf{r}_m^{(2)}, m = 1,\cdots,M-2$ that do not contain any contributions from terminals K-th and K-1. In general, after stage j, where j=1,2,..,K-1 contributions from terminals K,K-1,...,K-j+1 are canceled out and we are left with M-j modified received signal vectors $\mathbf{r}_m^{(j)}, m = 1,\cdots,M-j, j = 1,2,\cdots,K-1$ that contain signals due to terminals 1,2,...,K-j only. In this case, we will have

$$\mathbf{r}_m^{(j)} = \sum_{k=1}^{K-j} \mathbf{H}_{k,m}^{(j)} \cdot \mathbf{c}_k + \boldsymbol{\eta}_m^{(j)}, \quad m = 1,2,\cdots,M-j \tag{63}$$

where $\mathbf{H}_{k,m}^{(j)}$ and $\boldsymbol{\eta}_m^{(j)}$ are given by

$$\mathbf{r}_m^{(j)} = \mathbf{r}_m^{(j-1)} - \mathbf{H}_{K-j,m}^{(j-1)} \left(\mathbf{H}_{K-j,M-j}^{(j-1)}\right)^+ \mathbf{r}_{M-j}^{(j-1)} \qquad m = 1,2,\cdots M-j \tag{64}$$

$$\mathbf{H}_{k,m}^{(j)} = \mathbf{H}_{k,m}^{(j-1)} - \mathbf{H}_{K-j,m}^{(j-1)} \left(\mathbf{H}_{K-j,M-j}^{(j-1)}\right)^+ \mathbf{H}_{k,M-j}^{(j-1)}, \quad 1 \le m \le M-j, \, 1 \le k \le K-j \tag{65}$$

$$\boldsymbol{\eta}_m^{(j)} = \boldsymbol{\eta}_m^{(j-1)} - \mathbf{H}_{K-j,m}^{(j-1)} \left(\mathbf{H}_{K-j,M-j}^{(j-1)}\right)^+ \boldsymbol{\eta}_{M-j}^{(j-1)}, \quad m = 1,2,\cdots,M-j \tag{66}$$

This process is repeated until we are left with M-K+1 modified received signal vectors $\mathbf{r}_m^{(K-1)}, m = 1,\cdots,M-K+1$ that contain only contributions due to the first terminal. In this case we will have

$$\mathbf{r}_m^{(K-1)} = \mathbf{H}_{1,m}^{(K-1)} \cdot \mathbf{c}_1 + \boldsymbol{\eta}_m^{(K-1)}, \quad m = 1,2,\cdots,M-K+1 \tag{67}$$

which contains signals due to the first terminal only. Similarly, the modified channel matrix $\mathbf{H}_{1,m}^{(K-1)}$, $m = 1,2,\cdots,M-K+1$, will have a similar structure and is also orthogonal. Hence, it is straight forward to see that the ML decoder for signals from the first terminal is given by

$$\hat{\mathbf{c}}_1 = \arg\min_{\hat{\mathbf{c}}_1 \in C} \sum_{m=1}^{M-K+1} \left\| \mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)} \cdot \hat{\mathbf{c}}_1 \right\|^2 \tag{68}$$

and the corresponding uncertainty will be given by

$$\Delta_1 = \sum_{m=1}^{M-K+1} \left\| \mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)} \cdot \hat{\mathbf{c}}_1 \right\|^2 \tag{69}$$

Similarly, since the modified channel matrices $\mathbf{H}_{1,m}^{(K-1)}$, $1 \le m \le M-K+1$ are orthogonal, as before, the decoding rule in (68) will decompose into Q

separate rules for decoding $c_{11}, c_{12}, \cdots, c_{1Q}$. ***We may observe that the basic case for zero-forcing IC and ML decoding that we discussed in detail earlier is a special case of the above approach.***

$$
\begin{array}{l}
(\hat{\mathbf{c}}, \Delta) = \text{G\_ZFDECODE}\left(\{\mathbf{r}_m\}_{1\le m\le M}, \{\mathbf{H}_{km}\}_{1\le k\le K, 1\le m\le M}\right) \\
\{ \\
\quad \mathbf{r}_m^{(0)} = \mathbf{r}_m,\ 1 \le m \le M \\
\quad \mathbf{H}_{k,m}^{(0)} = \mathbf{H}_{k,m},\ 1 \le m \le M,\ 1 \le k \le K \\
\quad \text{for } j = 1 \to K-1 \\
\qquad M_j = M - j,\ K_j = K - j \\
\qquad \text{for } i = 1 \to M_j \\
\qquad\quad \mathbf{r}_i^{(j)} = \mathbf{r}_i^{(j-1)} - \mathbf{H}_{K_j i}^{(j-1)}\left(\mathbf{H}_{K_j, M_j}^{(j-1)}\right)^+ \mathbf{r}_{M_j}^{(j-1)} \\
\qquad\quad \mathbf{H}_{k,i}^{(j)} = \mathbf{H}_{k,i}^{(j-1)} - \mathbf{H}_{K_j, m}^{(j-1)}\left(\mathbf{H}_{K_j, M_j}^{(j-1)}\right)^+ \mathbf{H}_{k, M_j}^{(j-1)},\ 1 \le k \le K_j \\
\qquad \text{end} \\
\quad \text{end} \\
\quad \hat{\mathbf{c}} = \arg\min_{\hat{\mathbf{c}}_1 \in C} \sum_{m=1}^{M-K+1} \left\| \mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)} \cdot \hat{\mathbf{c}} \right\|^2 \\
\quad \Delta_1 = \sum_{m=1}^{M-K+1} \left\| \mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)} \cdot \hat{\mathbf{c}} \right\|^2 \\
\}
\end{array}
$$

The above-disclosed technique can be easily implemented within a detector 25 that comprises a stored program general purpose processor. Specifically, a subroutine $(\mathbf{c},\Delta)$=G_ZFDECODE($\{\mathbf{r}_m\}_{1\le m\le M}$, $\{\mathbf{H}_{km}\}_{1\le k\le K, 1\le m\le M}$) can be installed which returns the values $\mathbf{c},\Delta$ in response to submitted inputs $\{\mathbf{r}_m\}_{1\le m\le M}$ and $\{\mathbf{H}_{km}\}_{1\le k\le K, 1\le m\le M}$, as shown above.

Minimum Mean-Squared Error IC and ML Decoding Scheme: The MMSE IC and ML decoding in the general case can be developed in a similar fashion as follows. We recall the received signal vector at the m-th receive antenna in equation (57)

$$\mathbf{r}_m = \sum_{k=1}^{K} \mathbf{H}_{km} \cdot \mathbf{c}_k + \eta_m, \qquad m = 1,2,\cdots,M \tag{70}$$

This can be written in a matrix form as in equation (38)

$$\mathbf{r} = \mathbf{H} \cdot \widetilde{\mathbf{c}} + \mathbf{n} \tag{71}$$

where $\mathbf{r} = \begin{bmatrix} \mathbf{r}_1^T & \mathbf{r}_2^T & \cdots & \mathbf{r}_M^T \end{bmatrix}^T$ is a $ML\times 1$ vector, $\widetilde{\mathbf{c}} = \begin{bmatrix} \mathbf{c}_1^T & \mathbf{c}_2^T & \cdots & \mathbf{c}_K^T \end{bmatrix}^T$ is $QK\times 1$ a vector, $\mathbf{n} = \begin{bmatrix} \eta_1^T & \eta_2^T & \cdots & \eta_M^T \end{bmatrix}^T$ is a $ML\times 1$ vector, and

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_{11} & \mathbf{H}_{21} & \cdots & \mathbf{H}_{K1} \\ \mathbf{H}_{12} & \mathbf{H}_{22} & \cdots & \mathbf{H}_{K2} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{H}_{1M} & \mathbf{H}_{2M} & \cdots & \mathbf{H}_{KM} \end{bmatrix} \tag{72}$$

is the $ML\times QK$ channel matrix. As before, we redefine the vector $\mathbf{r}$ as $\mathbf{r} = \begin{bmatrix} r_1 & r_2 & \cdots & r_{ML} \end{bmatrix}^T$. As before, we assume that we are interested in decoding the symbols transmitted from terminal 1 $c_{11}, c_{12}, \cdots, c_{1Q}$. As before, when seeking to detect and decode signals $c_{11}, c_{12}, \cdots, c_{1Q}$ by minimizing a mean-squared error criterion, the goal is find a linear combination of the received signals such that the mean-squared error in detecting the signals $c_{11}, c_{12}, \cdots, c_{1Q}$ is minimized. In general terms, this can be expressed by an error cost function that is to be minimized, such as the function

$$J(\alpha,\beta) = \left\| \sum_{i=1}^{LM} \alpha_i r_i - \sum_{j=1}^{Q} \beta_j c_{1j} \right\|^2 = \left\| \alpha \cdot \mathbf{r} - \beta \cdot \mathbf{c}_1 \right\|^2 \tag{73}$$

Similarly, as before we can see that one of the $\beta_j, 1 \le j \le Q$ must be set to 1 or else we get an all zero solution for α and β. Consider the case where we set $\beta_j = 1$. Hence, in this case, the criteria to be minimized is

$$J_j(\alpha_j,\beta_j) = \left\| \sum_{i=1}^{LM+Q-1} \alpha_{ji} r_{ji} - c_{1j} \right\|^2 = \left\| \widetilde{\alpha}_j \widetilde{\mathbf{r}}_j - c_{1j} \right\|^2, \; 1 \le j \le Q \tag{74}$$

where

$$\widetilde{\alpha}_j = [\alpha_{j1}, \alpha_{j12}, \cdots, \alpha_{jLM}, -\beta_1, \cdots, -\beta_{j-1}, -\beta_{j+1}, \cdots, -\beta_Q] = \begin{bmatrix} \alpha_j & -\beta_j \end{bmatrix} \tag{75}$$

$$\tilde{\mathbf{r}}_j = \begin{bmatrix} \mathbf{r}_j^T & c_{11} & \cdots & c_{1j-1} & c_{1j+1} & \cdots & c_{1Q} \end{bmatrix}^T \qquad (76)$$

If we follow the same steps as in the basic case, we arrive at the conclusion that

$$\begin{aligned} \beta_i(j) &= 0 \quad i = 1, \cdots Q, \quad i \neq j \\ &= 1 \quad i = j \end{aligned} \qquad (77)$$

$$\alpha_j^* = \mathbf{M}^{-1}\mathbf{h}_j, \; 1 \leq j \leq Q \qquad (78)$$

where $\mathbf{h}_j$ is the j-th column of the channel matrix $\mathbf{H}$, and $\mathbf{M} = \mathbf{H}\mathbf{H}^* + \frac{1}{\Gamma}\mathbf{I}$, is an $ML \times ML$ matrix, Γ is the signal to noise ratio, and $\mathbf{I}$ is the $ML \times ML$ identity matrix.

In this case, as before, the error in decoding the j-th symbol c_{1j} will be minimized without any regard to the other symbols. Hence, the MMSE-IC and ML decoder will consist of Q different combiners, one for each symbol. It should be clear now that the MMSI-IC solution for the general case is a straight forward extension to the basic case shown earlier. The MMSE-IC solution for the general case can be implemented using the subroutine G_MMSE.DECODE shown below.

$$(\hat{\mathbf{c}},\Delta) = \mathrm{G\_MMSEDECODE}\left(\{\mathbf{r}_m\}_{1\le m\le M}, \{\mathbf{H}_{km}\}_{1\le k\le K, 1\le m\le M}, \Gamma\right)$$

{

$$\mathbf{r} = \begin{bmatrix}\mathbf{r}_1^T & \mathbf{r}_2^T & \cdots & \mathbf{r}_M^T\end{bmatrix}^T$$

$$\mathbf{H} = \begin{bmatrix}\mathbf{H}_{11} & \mathbf{H}_{21} & \cdots & \mathbf{H}_{K1} \\ \mathbf{H}_{12} & \mathbf{H}_{22} & \cdots & \mathbf{H}_{K2} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{H}_{1M} & \mathbf{H}_{2M} & \cdots & \mathbf{H}_{KM}\end{bmatrix}$$

$$\mathbf{M} = \mathbf{H}\mathbf{H}^* + \frac{1}{\Gamma}\mathbf{I}$$

for $j = 1 \to Q$

$\mathbf{h}_j = j$ - th column of $\mathbf{H}$

$$\alpha_j^* = \mathbf{M}^{-1}\mathbf{h}_j$$

$$c_j = \arg\min_{\hat{c}_j \in C} \left\|\alpha_j^* \mathbf{r} - \hat{c}_j\right\|^2$$

$$\Delta_j = \left\|\alpha_j^* \mathbf{r} - \hat{c}_j\right\|^2$$

end

$$\hat{\mathbf{c}} = \begin{bmatrix}c_1 & c_2 & \cdots & c_Q\end{bmatrix}^T$$

$$\Delta = \sum_{j=1}^{Q} \Delta_j$$

}

We claim:

1. A method for decoding a set of M signals received at an input interface from a plurality of terminal units that transmit on a given channel, comprising the steps of:

processing said M signals, received over L time intervals, where L is an integer, with signals having components related to channel coefficients between transmitting points of said terminal units and an input interface, to detect signals transmitted from each of said terminal units by canceling interference from K of said terminal units, where M and K are integers such that $M \geq 2$ and $K \leq M$, and to identity probable signals transmitted by said terminal units through maximum likelihood detection; and

applying signals developed through said maximum likelihood detection to a location, from which the signals may be applied to post processing that culminates in signals adapted for delivery to users.

2. The method of claim 1 where transmissions of said terminal units are synchronized.

3. The method of claim 1 where each of the K terminal units employs at least two transmitting antennas, and said input interface includes M receiving antennas.

4. The method of claim 1 where each of the K terminal units employs N transmitting antennas, where N is greater than 1, and said input interface includes M receiving antennas.

5. The method of claim 4, where said interference canceling involves algebraic operations carried out on said M received signals with said signals having components related to channel coefficients between transmitting points of said terminal units, to form a plurality of signals, each of which is

substantially unrelated to signals sent by all but one of the terminal units.

6. The method of claim 4 where said K=M=2 and $1 \le N \le 2$.

7. The method of claim 6 where said processing comprises multiplying said vector of received signals by conditioning matrix $\begin{bmatrix} \mathbf{I} & -\mathbf{G}_1\mathbf{G}_2^{-1} \\ -\mathbf{H}_2\mathbf{H}_1^{-1} & \mathbf{I} \end{bmatrix}$, where $\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^{*} & -h_{11}^{*} \end{bmatrix}$, $\mathbf{G}_1 = \begin{bmatrix} g_{11} & g_{12} \\ g_{12}^{*} & -g_{11}^{*} \end{bmatrix}$, $\mathbf{H}_2 = \begin{bmatrix} h_{21} & h_{22} \\ h_{22}^{*} & -h_{21}^{*} \end{bmatrix}$, $\mathbf{G}_2 = \begin{bmatrix} g_{21} & g_{22} \\ g_{22}^{*} & -g_{21}^{*} \end{bmatrix}$, **I** is the diagonal matrix, h_{ij} is a channel coefficient estimate between a transmitting antenna i of a first transmitting unit and a receive antenna j of said M receiving antennas, and g_{ij} is a channel coefficient estimate between a transmitting antenna i of a second transmitting unit and a receive antenna j of said M receiving antennas.

8. The method of claim 7 where L=2.

9. The method of claim 7 where said maximum likelihood detection minimizes a metric $\left\|\tilde{\mathbf{r}}_1 - \tilde{\mathbf{H}} \cdot \hat{\mathbf{c}}\right\|^2$ over all potential code vectors $\hat{\mathbf{c}}$, where $\tilde{\mathbf{r}}_1$ is one of said plurality of signals formed by said step of multiplying, and $\tilde{\mathbf{H}} = \mathbf{H}_1 - \mathbf{G}_1\mathbf{G}_2^{-1}\mathbf{H}_2$.

10. The method of claim 9 where said maximum likelihood detection develops an uncertainty measure.

11. The method of claim 10 where the uncertainty measure is given by $\Delta_c = \left\|\tilde{\mathbf{r}}_1 - \tilde{\mathbf{H}} \cdot \hat{\mathbf{c}}\right\|^2$.

12. The method of claim 6 where said processing executes a

subroutine $\mathrm{ZF.DECODE}(\mathbf{r}_x, \mathbf{r}_y, \mathbf{H}_x, \mathbf{H}_y, \mathbf{G}_x, \mathbf{G}_y)$ which returns an output $\hat{\mathbf{c}}$ by executing the following calculations

$$\tilde{\mathbf{r}} = \mathbf{r}_x - \mathbf{G}_x \mathbf{G}_y^{-1} \mathbf{r}_y,$$

$$\tilde{\mathbf{H}} = \mathbf{H}_x - \mathbf{G}_x \mathbf{G}_y^{-1} \mathbf{H}_y, \quad \text{and}$$

$$\hat{\mathbf{c}} = \arg\min_{\mathbf{c} \in \mathbf{C}} \left\| \tilde{\mathbf{r}} - \tilde{\mathbf{H}} \cdot \mathbf{c}_x \right\|^2,$$

where

$\mathbf{r}_x$ is the first argument of the ZF.DECODE subroutine and corresponds to a vector of signals received over said L time intervals at receiving antenna x,

$\mathbf{r}_y$ is the second argument of the ZF.DECODE subroutine and corresponds to a vector of signals received over said L time intervals at receiving antenna y,

$\mathbf{H}_x$ is the third argument of the ZF.DECODE subroutine and corresponds to a matrix comprising channel coefficients between transmitting antenna x of a first terminal unit and said M receiving antennas,

$\mathbf{H}_y$ is the fourth argument of the ZF.DECODE subroutine and corresponds to a matrix comprising channel coefficients between transmitting antenna y of said first terminal unit and said M receiving antennas,

$\mathbf{G}_x$ is the fifth argument of the ZF.DECODE subroutine and corresponds to a matrix comprising channel coefficients between transmitting antenna x of a second terminal unit and said M receiving antennas,

$\mathbf{G}_y$ is the sixth argument of the ZF.DECODE subroutine and corresponds to a matrix comprising channel coefficients between transmitting antenna y of said second terminal unit and said M receiving antennas, and

$\hat{\mathbf{c}}$ is a selected code vector $\mathbf{c}$ from a set of code vectors $\mathbf{C}$ as the code transmitted by the terminal related to the second and third arguments of the ZF.DECODE subroutine.

13. The method of claim 12 where said processing comprises

executing the ZF.DECODE subroutine a first time with argument $(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2)$, and executing the ZF.DECODE subroutine a second time with argument $(\mathbf{r}_2,\mathbf{r}_1,\mathbf{H}_2,\mathbf{H}_1,\mathbf{G}_2,\mathbf{G}_1)$.

14. The method of claim 12 where said subroutine ZF.DECODE also returns an uncertainty measure output Δ_c by carrying out the calculation $\Delta_c = \left\|\tilde{\mathbf{r}} - \tilde{\mathbf{H}}\cdot\hat{\mathbf{c}}\right\|^2$.

15. The method of claim 14 where said processing comprises the steps of:

executing the ZF.DECODE subroutine a first time, with argument $(\mathbf{r}_1,\mathbf{r}_2,\mathbf{H}_1,\mathbf{H}_2,\mathbf{G}_1,\mathbf{G}_2)$, to obtain a first version of a probable vector signal of a first terminal unit, $\hat{\mathbf{c}}_0$, and a measure of uncertainty, $\Delta_{c,0}$;

obtaining a first version of a probable vector signal of a second terminal unit, $\hat{\mathbf{s}}_0$, and a measure of uncertainty $\Delta_{s,0}$;

executing the ZF.DECODE subroutine a second time, with argument $(\mathbf{r}_2,\mathbf{r}_1,\mathbf{G}_2,\mathbf{G}_1,\mathbf{H}_2,\mathbf{H}_1)$, to obtain a second version of a probable vector signal of the second terminal unit, $\hat{\mathbf{s}}_0$, and a measure of uncertainty, $\Delta_{s,1}$;

obtaining a second version of a probable vector signal of a first terminal unit, $\hat{\mathbf{c}}_0$, and a measure of uncertainty $\Delta_{c,1}$;

selecting said first version of said probable vector signals if $(\Delta_{c,0} + \Delta_{s,0}) < (\Delta_{c,1} + \Delta_{s,1})$, and selecting said second version of said probable vector signal based otherwise.

16. The method of claim 1 where M > 2 and K > 2, and where said processing comprises the steps of:

combining said M received signals for form a signal that is essentially independent of transmissions of all but one of the terminal units,

performing a maximum likelihood detection to identify a probably

signal vector transmitted by said one of the terminal units,

modifying remaining ones of said M received signals to account for contribution of the probable signals vector, and

returning to said step of combining as long as there is a terminal unit whose transmission is to be detected, to select another one of said terminal units whose signal is to be detected with said step of maximum likelihood detection.

17. The method of claim 1 where M > 2 and K > 2, and where said processing comprises executing a subroutine

$$(\hat{\mathbf{c}},\Delta) = \mathrm{G\_ZFDECODE}\left(\{\mathbf{r}_m\}_{1\le m\le M},\{\mathbf{H}_{km}\}_{1\le k\le K,1\le m\le M}\right)$$

$$\{$$

$$\mathbf{r}_m^{(0)} = \mathbf{r}_m,\ 1\le m\le M$$

$$\mathbf{H}_{k,m}^{(0)} = \mathbf{H}_{k,m},\ 1\le m\le M,\ 1\le k\le K$$

$$\text{for } j = 1 \to K-1$$

$$M_j = M-j,\ K_j = K-j$$

$$\text{for } i = 1 \to M_j$$

$$\mathbf{r}_i^{(j)} = \mathbf{r}_i^{(j-1)} - \mathbf{H}_{K_j,i}^{(j-1)}\left(\mathbf{H}_{K_j,M_j}^{(j-1)}\right)^{+}\mathbf{r}_{M_j}^{(j-1)}$$

$$\mathbf{H}_{k,i}^{(j)} = \mathbf{H}_{k,i}^{(j-1)} - \mathbf{H}_{K_j,m}^{(j-1)}\left(\mathbf{H}_{K_j,M_j}^{(j-1)}\right)^{+}\mathbf{H}_{k,M_j}^{(j-1)},\quad 1\le k\le K_j$$

$$\text{end}$$

$$\text{end}$$

$$\hat{\mathbf{c}} = \arg\min_{\hat{\mathbf{c}}_i\in C}\sum_{m=1}^{M-K+1}\left\|\mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)}\cdot\hat{\mathbf{c}}\right\|^2$$

$$\Delta = \sum_{m=1}^{M-K+1}\left\|\mathbf{r}_m^{(K-1)} - \mathbf{H}_{1,m}^{(K-1)}\cdot\hat{\mathbf{c}}\right\|^2$$

$$\}$$

where

$\mathbf{r}_m^{(a)}$ is a received signal at antenna m, as modified in step (a);

$\mathbf{H}_{k,m}^{(a)}$ is a matrix of channel coefficients between terminal unit k and receive antenna m, at step a; and

$(\mathbf{H}_{k,m}^{(a)})^{+}$ is the generalized inverse of $\mathbf{H}_{k,m}^{(a)}$.

18. The method of claim 1 where said processing form a vector of signals from said M signals and said processing includes pre-multiplying said vector by a preconditioning matrix having components related to said channel coefficients.

19. The method of claim 18 where said components of the conditioning matrix are related to estimates of said channel coefficients.

20. A detector responsive to M received signals, where $M \geq 2$, from a plurality of terminal units and canceling interference from K of said terminal units, where $K \leq M$, comprising:

a processor for transforming said M received signals through algebraic operations involving components related to channel coefficients between transmitting antennas of said terminal units and said M received signal to form a plurality of signals, to each of which is substantially unrelated to signals sent by all but one of the terminal units, where M, N, and K are positive integers, and to perform maximum likelihood detection calculations on said plurality of signals; and

a memory coupled to said processor for storing information for controlling said processor.

21. A detector responsive to M received signals, where $M \geq 2$, from a plurality of terminal units and canceling interference from K of said terminal units, where $K \leq M$, comprising:

a first means responsive to said M received signals, for transforming said M received signals through algebraic operations involving components related to channel coefficients between transmitting antennas of said terminal units and said M received signal to form a plurality of signals, each of which is substantially unrelated to signals sent by all but one of the terminal units,

where M, N, and K are positive integers;

a second means, responsive to said first means, for performing maximum likelihood detection calculations on said plurality of signals; and

means for applying information developed by said second means to a location accessible for further processing and delivery to a user.

22. The detector of claim 20 where the processor also calculates an uncertainty measure associated with each maximum likelihood detection.

23. The detector of claim 20 where said processor forms said plurality of signals one at a time, and with each formed signal of said plurality of signals the processor selects a probable transmitted signal vector of one of said terminals and evaluates an uncertainty measure.

24. The detector of claim 20 where said processor, in the course of forming one of said signals from said M received signals, substantially nullifies contribution of transmissions from terminals for which a probably transmitted signal vector has been selected.

25. The detector of claim 20 where said processor:

forms a first version of a first of said plurality of signals, selects a first version of a probable transmitted signal vector of a first of said terminals, and evaluates a first uncertainty measure, then

forms a first version of a second of said plurality of signals and in the course of said forming said first version of said second of said plurality of signals nullifies contribution of transmissions of said first version of a probable transmitted signal vector of said first of said terminals, selects a first version of a probable transmitted signal vector of a second of said terminals, and evaluates a second uncertainty measure, then

forms a second version of said second of said plurality of signals, without said nullifying, selects a second version of a probable transmitted

signal vector of said second of said terminals, and evaluates a third uncertainty measure, then

forms a second version of said first of said plurality of signals and in the course of said forming said second version of said first of said plurality of signals nullifies contribution of transmissions of said second version of a probable transmitted signal vector of said second terminal, selects a second version of a probable transmitted signal vector of said first of said terminals, and evaluates a fourth uncertainty measure, then

forms a first combination involving said first and second uncertainty measures, forms a second combination involving said third and fourth uncertainty measures, and selects either said first version or said second version of the probable transmitted signal vector of said first of said terminals and the probable transmitted signal vector of said second of said terminals based on whether said first combination is greater than said second combination.

26. The detector of claim 20 where M=N=K=2.

27. The detector of claim 20 where M>2, and K>2.

28. The detector of claim 20 where M>2, K>2, and N>2**29.** The detector of claim 20 where the processor also estimates parameters of channel through which said M signals traverse.

30. The detector of claim 22 further comprising M antennas and amplification and demodulation circuitry interposed between said M antennas and said processor.

31. The detector of claim 20 further comprising a subroutine stored in said memory and executed by said processor, which subroutine performs said transforming of said M received signals.

32. The detector of claim 24 where each execution of said subroutine develops one of said signals that is substantially unrelated to signals sent by all but one of the terminal units.

33 The detector of claim 25 where said subroutine is executed not less than K times.

34. The detector of claim 20 further comprising a subroutine stored in said memory and executed by said processor, which subroutine performs said transforming of said M received signals, followed by a maximum likelihood detection.

35. The detector of claim 20 further comprising means for estimating channel parameters.

36. The detector of claim 20 where said processor forms said plurality of signals by forming a vector from said M received signals and pre-multiplying said vector by a conditioning matrix of coefficients that are related to channel coefficients between said transmitting antennas and an input interface.

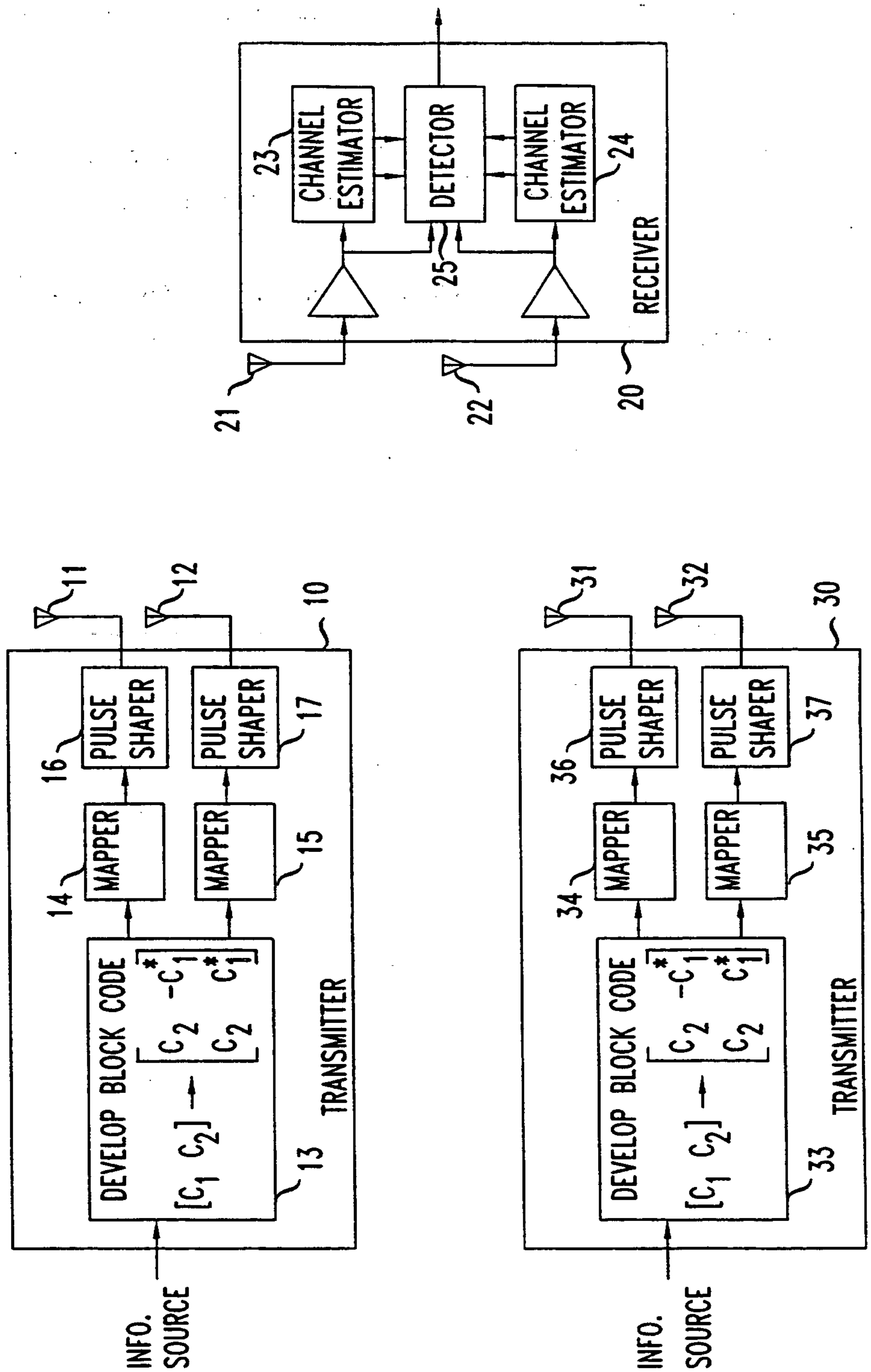
37. The detector of claim 13 where said conditioning matrix is $\begin{bmatrix} \mathbf{I} & -\mathbf{G}_1\mathbf{G}_2^{-1} \\ -\mathbf{H}_2\mathbf{H}_1^{-1} & \mathbf{I} \end{bmatrix}$, where $\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix}$, $\mathbf{G}_1 = \begin{bmatrix} g_{11} & g_{12} \\ g_{12}^* & -g_{11}^* \end{bmatrix}$, $\mathbf{H}_2 = \begin{bmatrix} h_{21} & h_{22} \\ h_{22}^* & -h_{21}^* \end{bmatrix}$, $\mathbf{G}_2 = \begin{bmatrix} g_{21} & g_{22} \\ g_{22}^* & -g_{21}^* \end{bmatrix}$, $\mathbf{I}$ is the diagonal matrix, h_{ij} is a channel coefficient estimate between a transmitting antenna i of a first transmitting unit and a receive antenna j of said M receiving antennas, and g_{ij} is a channel coefficient estimate between a transmitting antenna i of a

second transmitting unit and a receive antenna j of said M receiving antennas.

38. The detector of claim 37 where said processor also performs maximum likelihood detection on said plurality of signals.

39. The detector of claim 38, where said input interface comprises M receiving antennas.

FIG. 1


CHANNEL ESTIMATOR
DETECTOR
CHANNEL ESTIMATOR
RECEIVER
23
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PULSE SHAPER
PULSE SHAPER
MAPPER
MAPPER
DEVELOP BLOCK CODE
$[c_1\ c_2] \rightarrow \begin{bmatrix} c_2 & -c_1^* \\ c_2 & c_1^* \end{bmatrix}$
TRANSMITTER
INFO. SOURCE
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PULSE SHAPER
PULSE SHAPER
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MAPPER
DEVELOP BLOCK CODE
$[c_1\ c_2] \rightarrow \begin{bmatrix} c_2 & -c_1^* \\ c_2 & c_1^* \end{bmatrix}$
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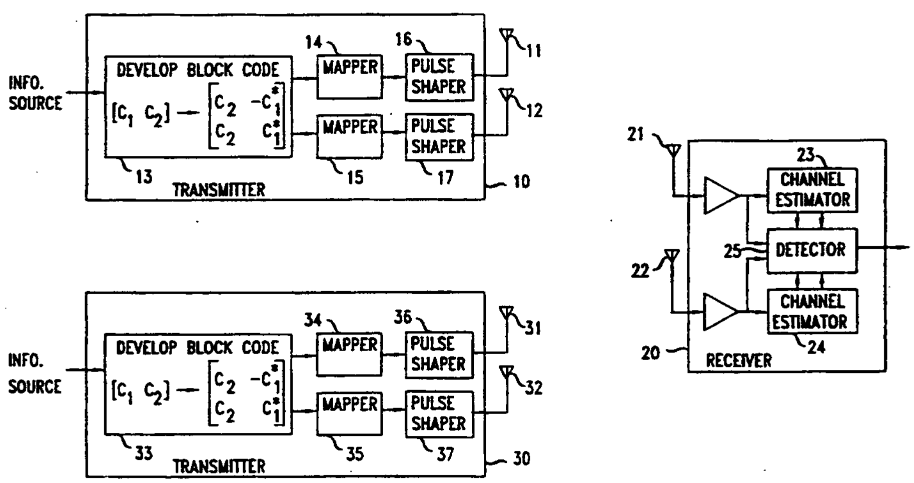

(57) Abstract

Block-encoded transmissions of a multi-antenna terminal unit are effectively detected in the presence of co-channel interfering transmissions from other multi-antenna terminal units, when the base station has a plurality of antennas, and interference cancellation is combined with maximum likehood decoding. The signals received in one base station antenna are employed in processing the signals received in a second base station antenna so as to cancel the signals of one terminal unit, while decoding the signals transmitted by the other terminal unit. Zero-forcing and MMSE approaches are presented. In another embodiment of the invention, the basic decoding approach is used to obtain an initial estimate for the symbols from each terminal. Assuming that the signals of the first terminal unit has been decoded correctly, the receiver employs this initial decoded signal of the first terminal unit to cancel their contribution to the signals received at the base station antennas while decoding the signals of the second terminal unit. This process is then repeated assuming that the signals of the second terminal unit have been decoded correctly; the receiver employs this initial decoded signal of the second terminal unit to cancel their contribution to the signals received at the base station antennas while decoding the signals of the first terminal unit. The above disclosed techniques are variable for any number K of terminal units concurrently transmitting over a given channel, where each terminal unit is using a space-time block code with N transmit antennas, and a base station has at least K receive antennas.

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INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 98/20982

A. CLASSIFICATION OF SUBJECT MATTER
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According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 4 057 758 A (HATTORI TAKESHI ET AL) 8 November 1977 see column 3, line 23 - column 3, line 65; claims 1,7; figures 1-5	1,3
A	PATENT ABSTRACTS OF JAPAN vol. 015, no. 351 (E-1108), 5 September 1991 & JP 03 135238 A (NIPPON TELEGR & TELEPH CORP), 10 June 1991 see abstract	1,3

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[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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INTERNATIONAL SEARCH REPORT

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C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
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